Comments* on "Some Notes on Strip Transmission Line and Waveguide Multiplexers"[1]

Use of two tuning screws through the ground planes of a strip-line cavity resonator can excite the parallel plane TEM mode unless both screws are equidistant from the center conductor, rendering this method of tuning quite unattractive. A single tuning screw parallel to and midway between the ground planes is the better way of tuning a stripline cavity resonator.

The implication that a Tchebycheff response shape is preferable to the Butterworth response shape merely on the sole criteria of sharper skirt selectivity is debatable. Butterworth filters are simpler to design, have more favorable phase responses, and are easier to align. Furthermore, Butterworth filters are superior when designing narrow band filters for minimum insertion loss.[2]

We are currently using direct-coupled waveguide resonant cavity filters and have found them to be quite satisfactory. These filters employ five cavity resonators with bandwidths of about 3 per cent in frequency (This corresponds to a filter Q of 33.) Use of quarter-wave coupled cavities would increase over-all filter length by a factor of two, while use of quarter-wave coupled resonant elements would result in intolerable filter insertion losses. It should also be noted that quarter-wave coupled waveguide filters usually employ nominal $\frac{3}{4}\lambda_g$ connecting lines, since use of $\lambda_g/4$ connecting lines and centered inductive coupling posts will cause appreciable higher-mode interaction between adjacent resonators. No general statement can be made concerning the relative merits of direct-coupled and quarter-wave coupled filters. Percentage bandwidth, physical size, construction cost, and other factors must be carefully considered in each specific filter application.

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[1] D. Alstadter and E. O. Houseman, Jr., 1958 WESCON CONVENTION RECORD, pt. 1, pp. 54–69.

[2] J. J. Taub and B. F. Bogner, "Design of three-resonator dissipative band-pass filters having minimum insertion loss," PROC. IRE, vol. 45, pp. 681–687; May, 1957.

The Representation of Impedances with Negative Real Parts in the Projective Chart*

In a previous note [1] the authors considered the representation of active networks in the reflection coefficient chart. The projective chart [2] is obtained from a stereographic projection onto a unit sphere and then from an orthographic projection from the sphere. It is immediately obvious that reflectances whose magnitude is greater than unity will lie within the unit circle of the projective chart. Therefore, each will coincide with another reflectance whose magnitude is less than unity.

From intuitive reasoning it is apparent the reflectances which share the same point in the projective chart are $re^{j\phi}$ and $(1/r)e^{j\phi}$. To show this analytically one determines the hyperbolic distance in the projective chart as proportional to the logarithm of the cross ratio. When the points[1] r and $1/r$ are transformed algebraically, the cross ratios are negatives (but both are real) and the hyperbolic distance of the reflectance whose magnitude is greater than unity is complex. The imaginary component of the distance arises because of the ultra-infinite end point of the measured distance. The real parts of both distances are equal, since the projective chart cannot indicate the j direction which is perpendicular to the plane of the projective chart (they are the same point). This may be seen by using the Riemann sphere as described by Bolinder [3] to visualize that the positive and negative resistance values are on different halves of the sphere and that the orthographic projection is perpendicular to the dividing equator and hence cannot differentiate between real conjugate impedances or reflectances which are inverse with respect to the unit circle. The equivalence between the inverse reflectances can also be seen by constructions in the plane by using the circle of inversion [4]. This leads to a modification of the β transformation of Deschamps which is shown in Fig. 1.

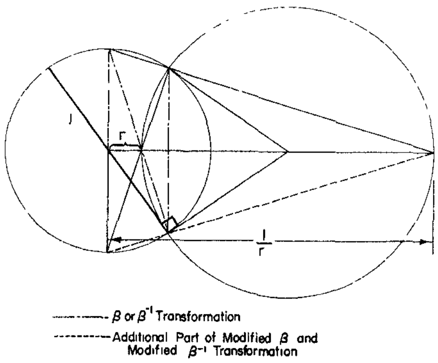

Fig. 1—Modified β and modified β^{-1} transformation.

Therefore, it is seen that the projective chart can be used to represent active networks by using the points of the real conjugate of the impedance or the reflectance that is inverse with respect to the unit circle.

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[1] L. J. Kaplan and D. J. R. Stock, "An extension of the reflection coefficient chart to include active networks," IRE TRANS. ON MICROWAVE THEORY AND TECHNIQUES, vol. MTT-7, pp. 298–299; April, 1959.
[2] G. A. Deschamps, "New chart for the solution of transmission-line and polarization problems," IRE TRANS. ON MICROWAVE THEORY AND TECHNIQUES, vol. MTT-1, pp. 5–13; March, 1953.
[3] E. F. Bolinder, "General method of analyzing bilateral two-part networks from three arbitrary impedance measurements," *Ericsson Technics*, vol. 14, no. 1, pp. 3–37; 1958.
[4] J. de Buhr, "Eine neue methode zur bearbeitung linearer vierpole," *FTZ-Fernmeldetech Z.*, vol. 8, pp. 200–204, April, 1955; pp. 335–340; June, 1955.

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[1] The phase angle ϕ will not affect the distance from the center.

Design Calculations for UHF Ferrite Circulators*

The recent advances in low-noise amplifier work for communications systems has created an additional demand for circulators; in this case, to prevent receiver noise from returning to the low-noise amplifier. In the range of frequencies greater than 2000 mc, ferrite circulators have been developed in circular and rectangular waveguides. However, in the UHF region, which is a range of frequencies of increasing interest and importance in communications, ferrite circulators present a problem in the sense that ordinary waveguides needed in this range are prohibitively large for practical use. Button[1] of Lincoln Laboratory and Seidel[2] of Bell Telephone Laboratories have pointed a way around this difficulty by considering a TEM structure (a coax) loaded antisymmetrically with dielectric material and ferrite. This configuration provides for the longitudinal component of RF magnetic field necessary for nonreciprocity in the phase constant.[2] The essentially TEM nature of the device allows use of reasonably small, practical cross-sectional areas. The parallel-plate analog analysis presented in Button's paper leads to a transcendental equation for the phase constant which we present below for convenience, together with an example of the structure (see Fig. 1).

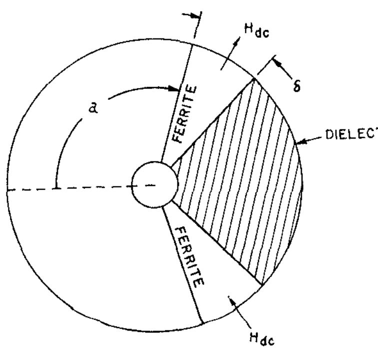

Fig. 1—Cross section of coaxial phase shifter.

* Received by the PGMTT, May 4, 1959.

[1] K. J. Button, *J. Appl. Phys.*, vol. 29, p. 998; June, 1958.

[2] H. Seidel, *J. Appl. Phys.*, vol. 28, p. 218; February, 1957.

$$L = 2(a + \delta) + \frac{2}{k_d} \tan^{-1}\left(\frac{-M}{N}\right) \qquad (1)$$

$$M = (\beta^2 G^2 - k_m^2 \rho^2) \cos k_a a \sin k_m \delta - k_a \sin k_a a [k_m \rho \cos k_m \delta + j\beta G \sin k_m \delta]$$

$$N = k_d [k_a \sin k_a a \sin k_m \delta - \cos k_a a (k_m \rho \cos k_m \delta - j\beta G \sin k_m \delta)]$$

$$G = \rho/\theta$$

$$\rho = \frac{\mu}{\mu^2 - \kappa^2}$$

$$\theta = -j\mu/\kappa$$

$$\mu = 1 + \frac{4\pi M_S \gamma f_0}{f_0^2 - f^2}; \quad \kappa = \frac{4\pi M_S \gamma f}{f_0^2 - f^2};$$

$$f_0 = \gamma H_{dc}$$

H_{dc} = static magentic field applied to ferrite
$\gamma = 2.8 \times 10^6$ cps-oersted
f_0 = ferrite resonant frequency
f = RF frequency
$4\pi M_S$ = saturation magnetization of the ferrite
δ = ferrite slab thickness
$2a$ = width of empty region
L = mean value of circumference of inner and outer conductors of coax

$$k_d^2 = \frac{\omega^2}{c^2} K_e - \beta^2$$

$\omega = 2\pi f$
c = speed of light in vacuum
K_e = dielectric constant of dielectric slab
β = phase constant

$$k_a^2 = \frac{\omega^2}{c^2} - \beta^2$$

$$k_m^2 = \frac{\omega^2}{c^2} \frac{K_f}{\rho} - \beta^2$$

K_f = dielectric constant of ferrite

We carried out computer solutions of the transcendental equation (1) for the differential phase shift $\beta_+ - \beta_-$ in the frequency range 400–440 mc. Our operating conditions were chosen in the following way. Calculations were made by Button[1] on a 2000-mc coaxial ferrite phase shifter. We scaled,[3] in a very approximate way, his operating conditions down to 400 mc. Thus we chose a ferrite slab thickness $\delta = 2$ cm (corresponding to Button's choice 0.425 cm), an empty space region $2a = 26$ cm vs Button's 5.7 cm, a ferrite magnetization 450 gauss vs Button's 800 gauss (procurement of required 200-gauss ferrite is not presently possible, so we chsse the lowest reasonable value for $4\pi M_S$), and a dc magnetic field of 100 oersteds vs Button's 275 oersteds. We chose a large dielectric constant $K_e = 15$ in order to attain large nonreciprocal phase shift. The dielectric constant of the ferrite was taken to be 10. Eq. (1) was then solved for $\beta_+ - \beta_-$ as a function of L for three sets of a, δ: $a = 15$ cm, $\delta = 2$ cm; $a = 15$ cm, $\delta = 1.6$ cm; $a = 13$ cm, $\delta = 2$ cm. The results are shown in Fig. 2 at 400 mc for $H_0 = 100$ oersteds. Practical coax is designed with a mean circumference of 33.9 cm and for this value, Fig. 2 shows the largest differential phase shift will occur for the set $a = 13$ cm, $\delta = 2$ cm (even larger $\Delta\beta$ will occur for still smaller a).

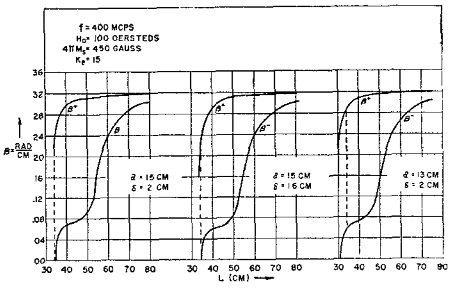

Fig. 2.

Keeping $a = 13$ cm, $\delta = 2$ cm, we then varied the dc field H_0 between 50 and 152 oersteds and also varied f between 400 and 440 mc and solved (1) for $\beta_+ - \beta_-$ for these varying conditions. Fig. 3 shows typical results, β_+ and β_- vs L, when $H_0 = 145$ oersteds and f is varied. This graph yields $\Delta\beta$ vs f for a given L (33.9 cm) and a given H_0. Allowing H_0 to take on different values then leads to Fig. 4, which shows $\Delta\beta$ vs f for 3 dc fields. From this figure, we see that best broadbanding is obtained for a dc field $H_0 = 145$ oersteds. The $\Delta\beta$ attained for these conditions is about 0.250 radian/cm leading to a circulator ($l\Delta\beta = \pi$ radians; l = length of circulator) about 5 inches long. The variation in $\Delta\beta$ in the 400 to 440-mc range for the 145-oersted field is about ± 1.3 per cent. This degree of broadbanding in $\Delta\beta$ leads to a minimum of 33-db isolation between circulator arms.

To summarize, Button's theory for the coax configuration when applied to the 400- to 440-mc band leads to the following design parameters (see Fig. 1):

ferrite slab thickness $\delta = 2$ cm
air region $= 2a = 26$ cm
ferrite magnetization = 450 gauss
dc magnetic field = 145 oersteds
mean circumference of coax = 33.9 cm
dielectric constant = 15
length of coax = 5 inches
isolation between arms ≥ 33 db over the band

It should be mentioned that 1) a more exhaustive investigation of the a and δ parameters might lead to larger $\Delta\beta$ and therefore even more compact structure, 2) the above design parameters can serve as the basis for design of coax circulators in other frequency bands by scaling.

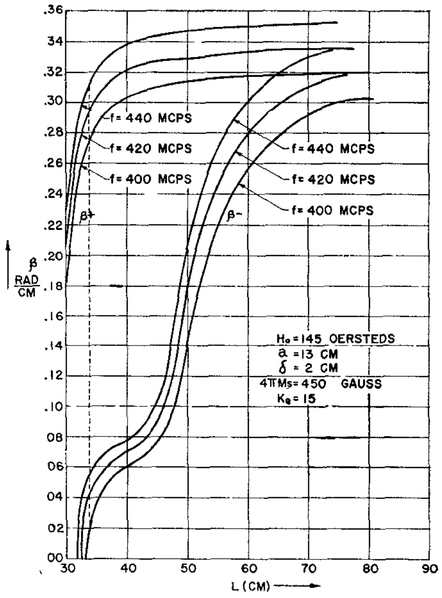

Fig. 3.

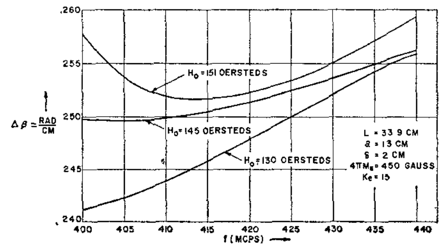

Fig. 4.

It also should be mentioned that the theory used here predicts nothing about the insertion loss of the device. However, we may note that the resonance condition for this structure [obtained by letting $\beta \to \infty$ in the transcendental equation (1)] is $\mu + \kappa = -1$ or $f_{resonance} = \gamma(H_{dc} + 2\pi M)$. Thus for $H_0 = 145$ oersteds, $2\pi M = 225$ gauss, and

$$\gamma = 2.8 \times 10^6 \frac{\text{cycles}}{\text{second oersted}},$$

we have $f_{resonce} \approx 1000$ mc, well above the operating range 400 to 440 mc. Thus, the device should have reasonably small insertion loss.

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[3] S. Weisbaum and H. Seidel, *Bell Sys. Tech J.*, vol. 35, p. 877; July, 1956.